A VMOS FET POWER AMPLIFIER

Charles Fletcher was first licensed at the age of 16 in 1948, following training as a Merchant Navy radio officer. After a short spell at sea he undertook further education, and has worked on navigational electronics (Decca), control and protection in electricity generation (CEGB), and latterly medical electronics and computing (Lincoln hospitals). His prime interest in amateur radio is to obtain top performance from low-cost homemade equipment, and his preferred communication mode is cw.

C F Fletcher, BSc, MIEE, CEng, G3DXZ*

THIS ARTICLE describes a cheap, medium-power, untuned high-gain amplifier intended as an add-on to QRP equipment or, alternatively, to form part of homebrew transmitter-receiver. The design has an air-proven performance record at G3DXZ and is well mannered and tolerant of abuse.

VMOS devices

The vmos power fet has been around now for several years, and despite its excellent qualities of ruggedness, stability and good high-frequency performance it is only commonly seen in its switching form. The mystery of why the rf devices are so rare is simply solved by asking the price; switching types £1 to £5, rf types £25 to £100! In this latter financial area, "experiment" becomes a dirty word! However, as the general characteristics of rf and switching types are generally similar, I was tempted to see what could be achieved using the low-cost switching types, goaded on by the old maxim that what anyone can do for £1, a good engineer can do for 25p. All very well, but a scrutiny of the published data [1] revealed the Achilles heel of the switching mosfet, ie rather high input and output capacitance. Of all the many devices listed, the only types offering real promise as a broadband amplifier were the VN90AA (2A 90V TO3 metal package) and its relations like VN88AF (2A 80V TO 202 plastic package). I used the VN90AA, available cheaply from [3], although the plastics-encapsulated variety also works well with slightly lower device dissipation.

Fig 1. (a) Schematic symbol of the standard unprotected vmos device. (b) The protective gate diode found in many switching devices

Before proceeding to discuss the circuitry, a few words regarding the characteristics of the vmos devices might be worthwhile. To anyone brought up in valve technology the mosfet is the nearest approach to a power pentode yet devised. For those of less advanced years, it is a big brother to the insulated gate-mosfets that have been around for years in small-signal guise.

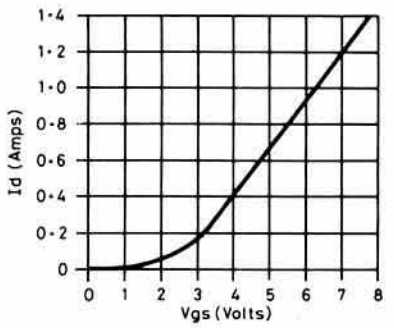


Fig 2. Input/output characteristic of the vmos device

Fig 1(a) shows the schematic symbol of the standard unprotected device and Fig 1(b) illustrates the protective gate diode found in many switching devices. The input/output characteristic is shown Fig 2. Note that conduction starts only when the gate is positive with respect to source, zero gate-source volts turns the fet firmly off. The gate can swing negative with the unprotected type, but negative excursions are limited to 0.3V for gate-protected devices. This latter limitation is the reason why normal tuned or transformer-coupled circuits cannot be applied directly to drive the gate protected mosfet. The following circuitry was developed to overcome this limitation and can be used with protected or unprotected devices. Finally, the family of devices chosen, which have tolerable input capacitances, unfortunately have an "on" resistance of 2Ω when driven hard. This relatively high drain-source resistance in a 2A device reduces their efficiency in low voltage applications, ie 12V ht use, and to make them effective as a power amplifier one must use the maximum drain-source voltage they will stand. Luckily these mosfets will work happily right up to their maximum voltage rating, which is one reason why they are referred to as rugged!

Driver section

The driver circuit is based on the classical long-tailed pair, and just in case some are not familiar with this old workhorse, Fig 3 shows the basic arrangement. The transistors, pnp in this case, have their emitters commoned, and dc bias voltages are developed by the Rb1, Rb2 potentiometers. The voltage across Rb1 less the emitter-base voltage of the transistors, which is fairly constant around 0·4V, appears across Rt, the so-called "tail" resistor. The current in Rt splits equally between TR1 and TR2 provided the base voltages at A and B are equal. In balance, therefore, the currents through and voltages across the collector load resistors Rc1 and Rc2 will be equal. Any voltage difference between A and B causes the balance to be disturbed and an amplified difference appears at C and D (**NB:** the changes at C and D have opposite polarity). The dc conditions are extremely stable, as the resistors dominate the circuit provided that Rt is not too small, or, in the language of the circuit engineer, provided the tail is long! Thus the circuit lends itself well to use as a driving stage for a push-pull vmos amplifier, allowing easy control of both ac and dc conditions.

Now refer to Fig 4, the actual driver circuit. Here Darlington-coupled pairs of transistors are used in place of the single devices used in the prototype circuit to give increased current gain to the amplifier and reduce the load on the rf drive transformer and bias network. The amplifier is brought into life when the tail current is switched on by TR2, the keyed line being taken to ground. Keyswitch current is 5mA. TR1 establishes the dc bias voltage for the long-tailed pair, and RV1 should be set to produce 3V dc across R7 and R8 with RV2 adjusted to maximum resistance (ie minimum gain condition). This is a convenient gate voltage for the vmos

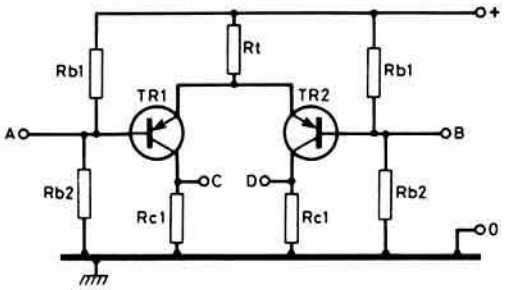


Fig 3. The long-tailed pair

*12 Park Crescent, Retford, Notts DN22 6UF.

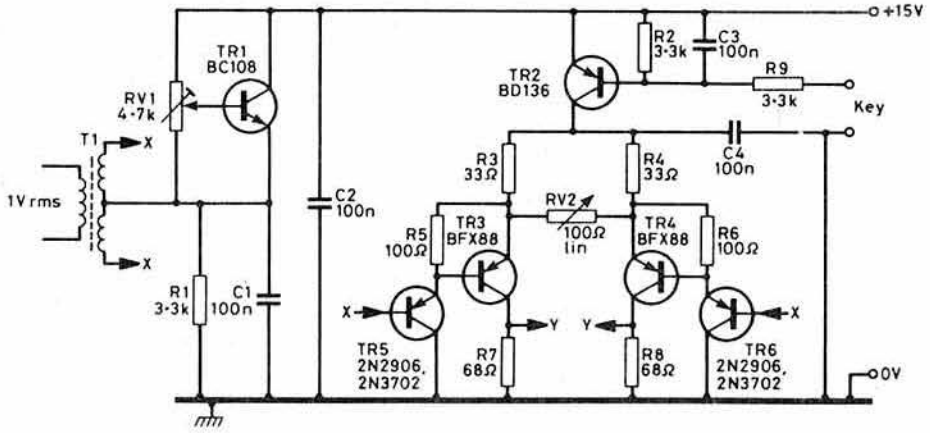


Fig 4. The driver circuit

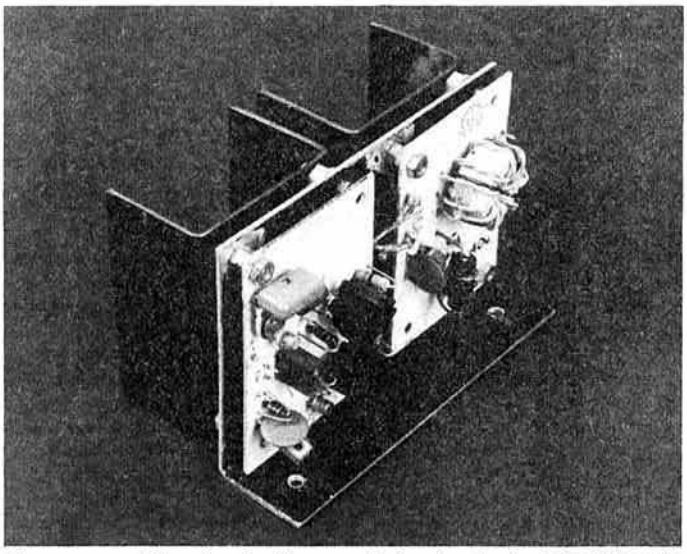
devices and puts the push-pull fets into Class AB1—old valve men take note! The resulting vmos standing current should be 500–600mA. The position of the rf drive control RV2 has little effect upon dc conditions but affects the rf gain as follows. Resistors R3 and R4 are introduced into the emitter circuits purposely to reduce the stage gain of the circuit. As RV2 is reduced it progressively shorts out R3 and R4 and brings the gain up to maximum. The rf drive is applied via T1 which is formed on a twin-hole ferrite bead. The centre-tapped secondary winding (wound bifilar) produces the push-pull rf drive voltage needed by the long-tailed pair from a single-ended source of 1V rms and low impedance (50Ω is adequate). The drive can be left on continuously, as when the key is up the amplifier is absolutely dead. This aspect of the design is a great help in using full QSK keying in dc transceivers, for which the amplifier was originally designed.

Power section

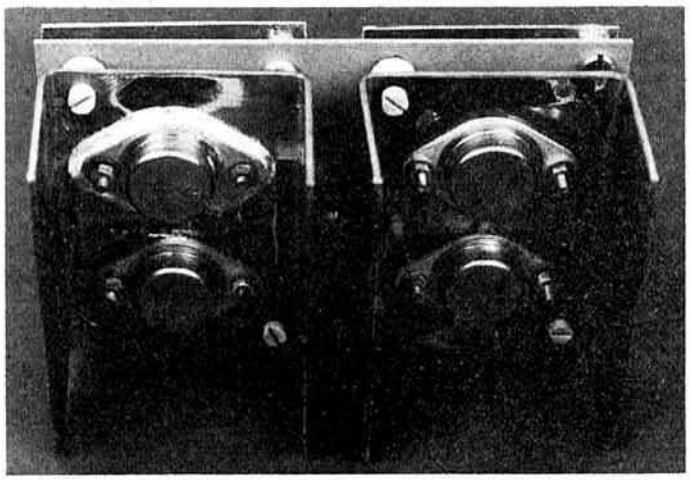
To achieve a dc input of 60W, the drain load resistance presented to the vmos pairs needs to be about 16Ω. This value is derived by the following reasoning. Assume an ht supply of 40V and that the vmos drain voltage will not swing below 4V. This assumption allows a peak drain voltage swing of 36V. For a dc input of 60W, the mean supply current must be 60/40 = 1·5A mean or 1·414 × 1·5 = 2·12A peak (assuming sinusoidal waveforms). So the drain load resistance to drop 36V at 2·12A must be 36/2·12 = 16·98Ω. Rounding down to 16Ω is permissible, as the vmos have plenty in hand and working backward through the logic will show that 16Ω allows 63·6W input. In practice, if the pa is hard driven and the waveform tends to become square, the input will be even higher. This latter condition is highly efficient but does produce more harmonic power; so beware if interference is a prime consideration.

Coupling the vmos to the atu and achieving a good impedance match together with adequate harmonic attenuation can be done in a number of ways. The two methods considered were (*a*) step-down impedance transformation balanced to unbalanced and an "L" matching section, and (*b*) step-up transformation and elliptic filter. Two good references on these topics are [4] and [5]. Given that a tuning range of less than 100kHz is adequate on the 3·5MHz band, the "L" section's simplicity wins, but if the full band needs to be covered without tuning—or use on 1·8MHz is considered—then the elliptic filter technique is superior. For exclusively 3·5MHz cw use, I chose the "L" section.

Push-pull amplifiers in good balance suppress even-order harmonics, leaving only the odd-order third, fifth and seventh to be controlled. An impedance-matching "L" section becomes effective as a filter when the input and output impedances are markedly different, ie when the effective "Q" is high. To achieve this, transformer T2 is wound to step down the impedance from 16Ω at the drain of the fet to 4Ω at the input of the "L" section, which in turn raises the impedance level to 50Ω suitable for a coaxial cable. The 4 to 50Ω impedance transformation achieved by the "L" section permits a reasonable operating "Q" and effectively removes the unwanted



The power amplifier, showing the separate input and output circuit boards



The parallel push-pull vmos fets (VN90AA) mounted on simple heatsinks run only warm to touch at 70W input

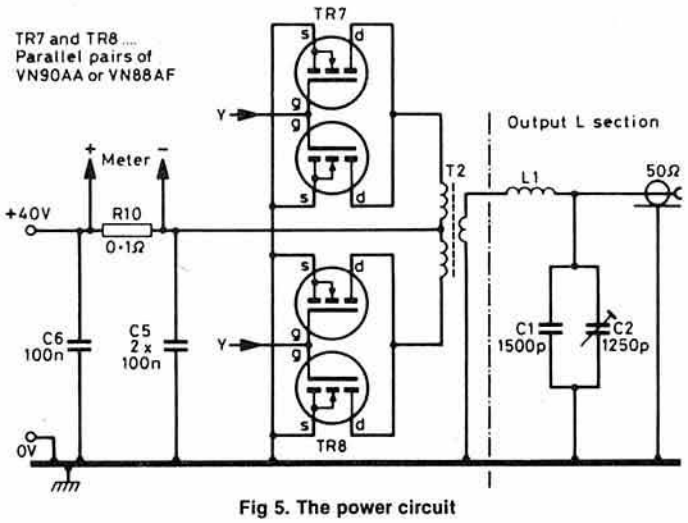


Fig 5. The power circuit

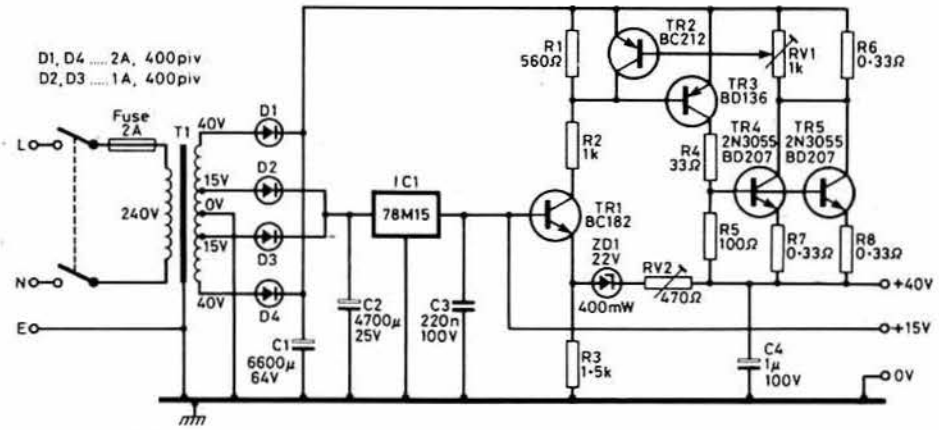


Fig 6. Regulated psu circuit

harmonics. I did not have a spectrum analyser but on examination with a good oscilloscope the rf output across a 50Ω resistive load was found to be remarkably pure. Operation in close proximity to audio and tv sets without any bad effect confirmed the clear output.

Use of a single toroidal transformer, type T106/2 core from [6], was found satisfactory for use on 3·5MHz, but should operation be attempted on higher frequencies a balancing transformer should be added (see almost any other broadband design). As well as the recommended iron-dust toroid, a wide variety of surplus cores were tried and almost all worked well on 3·5MHz although their performance was markedly poorer on 7 and 10MHz. The vmos fet is very tolerant of poor output transformers, only the output power suffers if the wrong type is used, leaving plenty of room for experiment.

Constructional notes

All components, excepting the vmos fets, are best mounted on small pcbs, leaving as much unused copper on the boards as possible and earthing it. I used two boards, one holding the driving side components and the other the output transformer and its associated capacitors and meter shunt. As always in rf circuitry, lead length should be as short as possible and the input separated from the output by earthed copper if it can be arranged. The leads to the 100Ω drive control rheostat must be short to avoid instability at high frequencies (ie above the operating frequency). Remember when laying out components on the boards that it is a balanced circuit and physical symmetry is good practice. The meter shunt R10 was made up to suit the meter available by winding a few turns of resistance wire on a 22Ω 2W carbon resistor.

The fets themselves should be mounted on separate heatsinks each around 2 by 3in and dish shaped for convenience. Using two heatsinks keeps the capacitance across the output transformer to a minimum and improves the higher frequency performance. The heatsinks can be supported on 0·25in insulating bushes or nylon screws, and a good method of construction is to bolt the heatsinks up to one side of an aluminium support screen, and the pcbs to the other side. This way, the strong rf fields on the heatsinks are kept away from the driving circuits. Again keep the gate and drain leads short and separate.

The coils in the output matching "L" sections were wound with 1·5mm plastic-insulated copper—mains wiring cable is good. Adjust inductance by stretching the coils. Mica compression trimmers are cheap, efficient and work well in hgh capacitance matching networks.

It is well to remember that in low impedance circuits handling 30 to 40W of rf energy, that circulating currents reach substantial proportions. For instance, 40W in a 4Ω circuit requires near 3·5A rms. This current is mainly carried near the surface of conductors, so stout wiring is repaid by low losses. Some suggested matching section data is given in the components list.

Components list

POWER AMPLIFIER

R1,2,9	3·3KΩ 0·25W
R3,4	33Ω 0·25W
R5,6	100Ω 0·25W
R7,8	68Ω 0·5W
R10	0·1Ω 1W (resistance wire wound on 2W resistor)
RV1	4·7kΩ skeleton preset
RV2	100Ω linear
C1,2,3,4,6	100nF 63V disc ceramic
C5	2 × 100nF 63V disc ceramic in parallel
T1	0·5 by 0·5 by 0·25in twin-hole bead. Pri 25t, sec 20-0-20t bifilar wound
T2	1 by 0·5in ferrite ring core. See text. Pri 20-0-20t 18swg bifilar wound and distributed around core, sec 10t 18swg distributed
TR1	BC108
TR2	BD136
TR3,4	BFX88
TR5,6	2N2906 or 2N3702
TR7,8	Parallel pairs VN90AA or VN88AF

OUTPUT L SECTION

L1 wound in 1·5mm plastic insulated mains wiring cable

3·5MHz	L1	5t 1in dia close-wound
	C1	1,500pF silvered mica 250V
	C2	1,250pF mica compression trimmer

POWER UNIT

C1	6,600µF 64V
C2	4,700µF 25V
C3	220nF 100V
C4	1µF 100V
D1,4	2A 400 piv
D2,3	1A 400 piv
R1	560 0·25W
R2	1kΩ 0·25W
R3	1·5kΩ 0·25W
R4	33Ω 0·25W
R5	100Ω 0·25W
R6,7,8	0·33Ω 2W
RV1	1kΩ skeleton preset
RV2	470 skeleton preset
IC1	78M15
T1	40-0-40, tapped 15-0-15, 3A
TR1	BC182
TR2	BC212
TR3	BD136
TR4,5	2N3055 or BD207
ZD1	22V 400mW zener diode

Power supply

A good power source for the vmos pa should have a stable output voltage under varying load conditions and be protected against overload. Good regulation of the output voltage, ie the amount the output voltage rises when the load is removed, is particularly important when using the fets near to their maximum voltage rating. My circuit is shown in Fig 6.

The driver section of the PA needs 15V, and this is easily produced using a 7815-type voltage regulator chip, 1A rating. The unregulated supply to the 7815 is not critical provided it does not fall below 18V at any time. Using too high an unregulated supply voltage produces unnecessary heat, and an unloaded voltage of 21 to 25V is a reasonable range which requires an ac transformer tap of 15 to 17V.

The 40V supply (assuming 80V fets) is a regulator circuit which uses the 7815 output as a reference. Over-current protection is provided by TR2, and RV1 sets the maximum current that can be drawn. The output voltage can be finely adjusted by RV2 which allows for some variation in the zener diode voltage. The design is simple and rugged. The transformer used in the prototype was a battery-charging type which had convenient taps. Again to reduce the excess heat generated by the pass transistors TR4 and TR5, the transformer should have an ac output between 38 and 42V. Using a finned heatsink rated at 2° C/W or less allows the 7815 and the pass transistors to be mounted together, and this is most easly done if plastic encapsulated devices are used. **NB:** If only 40W is desired, TR5 can be omitted.

References

[1] *VMOS Power FET Design catalogue*. Siliconix.
[2] *ARRL Handbook*, Chapter 6.
[3] J Birkett, The Strait, Lincoln.
[4] "An introduction to elliptical filters", J Williamson. *Rad Com* February 1983.
[5] "Design of L-Networks, J A Ewen. *Rad Com* August 1984.
[6] TMP Electronic Supplies, Pinfold Lane, Buckley, Clwyd CH7 3PL. □